Appln. No.: 10/536,571
Amendment Dated May 15, 2007
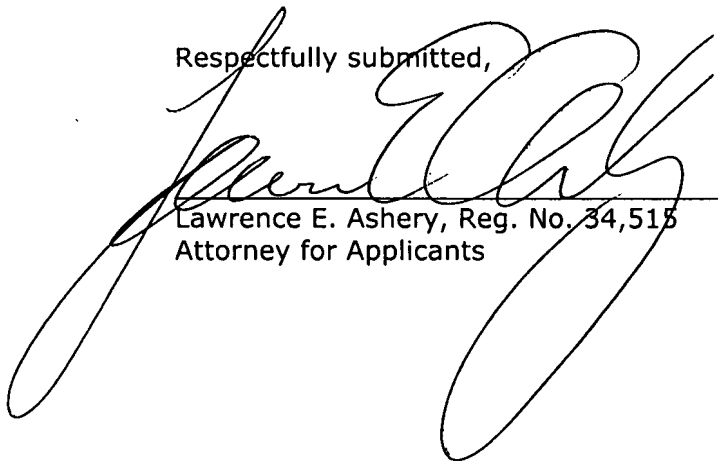Reply to Office Action of March 21, 2007

AOY-3992US

Remarks/Arguments:

The application has been indicated as being in condition for allowance except for formal matters. The Official Action required reference characters to be in parenthesis, required a modification to the characters used to list method steps, and objected to claim 18. In response, at substitute specification is attached, wherein characters have been placed in parenthesis, the claims have been renumbered, and claim 18 has been cancelled.

In view of the amendments set forth above, the above-identified is in condition for allowance, which action is respectfully requested.

Respectfully submitted,

Lawrence E. Ashery, Reg. No. 34,515
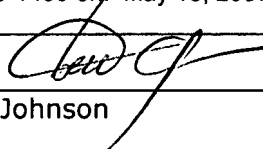Attorney for Applicants

LEA/bj

Dated: May 15, 2007

P.O. Box 980
Valley Forge, PA 19482
(610) 407-0700

The Director is hereby authorized to charge or credit Deposit Account No. **18-0350** for any additional fees, or any underpayment or credit for overpayment in connection herewith.

I hereby certify that this correspondence is being deposited with the United States Postal Service as first class mail, with sufficient postage, in an envelope addressed to: Commissioner for Patents, P.O. Box 1450, Alexandria, VA 22313-1450 on May 15, 2007.

Beth Johnson

149784

DESCRIPTION

METHOD AND APPARATUS FOR ESTIMATING ROTOR POSITION OF

SWITCHED RELUCTANCE MOTOR, AND METHOD AND APPARATUS FOR

5   SENSORLESS CONTROL OF SWITCHED RELUCTANCE MOTOR

This application is a U.S. National Phase application of PCT International Application No. PCT/JP2002/012412.

10

Technical field

The present invention relates to the development of closed loop control techniques for switched reluctance motors (SRMs) without a shaft position sensor.

15

Background art

A switched reluctance motor (SRM) is energized phase by phase in sequence to generate reluctance torque and enable smooth motor rotation. A schematic diagram of a

20   three phase switched reluctance motor is shown in Fig. 1. The number of strokes (N) in SRM per one mechanical revolution is dependent on the number of phases (M) and the number of rotor poles (P) and is given by,

$$N = M*P \quad (1).$$

25   Therefore, the stroke angle(S) in mechanical degrees is defined as,

$$S = 360° / N \quad (2).$$

When the number of poles is very large and the stroke angle is very small, the SRM is typically operated

30   in open loop as a variable reluctance stepper motor and

needs no knowledge of rotor position information during running condition.   On the other hand, when the number of poles is small and the stroke angle is very large, the SRM is generally operated in closed loop during running condition and hence, the knowledge of accurate rotor position information is very important to rotate the motor.

Accurate rotor position information is typically obtained from a shaft position sensor. Shaft position sensors are expensive and have reliability problems and hence, the sensorless operation of SRM is desired.

In a typical position sensorless operation of SRM, rotor position estimation is carried out either discretely i.e. once per stroke angle or continuously. Discrete rotor position estimation is ideal for applications where slow speed response is required where as, continuous rotor position estimation is carried out in applications where fast speed response is desired. Rotor position estimation of the SRM can be carried out from the pre-determined knowledge of its non-linear per phase flux-linkage/current characteristics or the inductance/current characteristics as shown in Fig. 2 and Fig. 3 respectively. Several non-linear analytical models based on the flux-linkage and the inductance characteristics of the motor with respect to the rotor position have been proposed [S.Saha, K.Ochiai, T.Kosaka, N.Matsui and Y.Takeda "Developing a sensorless approach for switched reluctance motors from a new analytical model", 1999 IEEE-IAS Annual meeting, Vol. 1, pp 525-532., and G.Suresh, B.Fahimi, K.M.Rahaman and M.Eshani "Inductance based position encoding for sensorless SRM drives", 1999 IEEE-PESC Annual Meeting, pp 832-837.] to

calculate the exact rotor position after sensing the d.c. link voltage V_{dc} and the phase current I_{ph} of the inverter drive circuit as shown in Fig. 4. A voltage sensor VS is used to sense the d.c. link voltage and three current

5 sensors CS are used to sense the current flowing in each phase winding 1u, 1v or 1w of the SRM 1. The traditional lock and forced drive method with open loop operation can be applied for the starting of these motors at no load. Rotor position estimation scheme is enabled at a minimum

10 speed (ω_{min}) which is normally 10% of the rated speed or higher. The motor moves from open loop to closed loop operation after the rotor position estimation scheme is enabled and then subsequently load is applied to the motor. The entire closed loop control scheme with the rotor

15 position estimation is implemented with the help of a micro-controller or a digital-signal processor.

Disclosure of invention

In a first aspect of the invention, provided is a

20 discrete rotor position estimation method for a synchronized reluctance motor. According to this estimation method, d.c.-link voltage V_{dc} and a phase current I_{ph} are sensed. A flux-linkage λ_{ph} of an active phase is calculated from the sensed d.c.-link voltage V_{dc}

25 and the sensed phase current I_{ph}. The calculated flux-linkage λ_{ph} is compared with a reference flux-linkage λ_r. The reference flux-linkage λ_r corresponds to a reference angle θ_r which lies between angles corresponding to aligned rotor position and non-aligned rotor position in

30 the synchronized reluctance motor. An estimated rotor

position θ_{cal} is obtained only once based on the comparison result when the calculated flux-linkage λ_{ph} is greater than the reference flux-linkage λ_r. From the estimated rotor position θ_{cal}, incremental rotor angle $\Delta\theta$ for every PWM

5   interrupt is calculated. The knowledge of $\Delta\theta$ is finally used to control the motor.

In a second aspect of the invention, provided is a discrete rotor position estimation method for a synchronized reluctance motor. According to the method in

10   the second aspect, a d.c.-link voltage V_{dc} and a phase current I_{ph} are sensed. A flux-linkage λ_{ph} of an active phase is calculated from the sensed d.c.-link voltage V_{dc} and the sensed phase current I_{ph}. The calculated flux-linkage λ_{ph} is compared with either two or three reference

15   flux-linkages such as $\lambda_{r1}, \ldots$. The reference flux-linkages $\lambda_{r1}, \ldots$ correspond to a reference rotor angles $\theta_{r1}, \ldots$ all of them lying between angles corresponding to aligned rotor position and non-aligned rotor position in the synchronized reluctance motor. Rotor positions $\theta_{cal1}, \ldots$ are obtained

20   based on the comparison results, twice or thrice when the calculated flux-linkage λ_{ph} is greater than the reference flux-linkage $\lambda_{r1}, \ldots$. From the estimated rotor positions $\theta_{cal1}, \ldots$ incremental rotor angles for every PWM interrupt are also calculated twice or thrice such as $\Delta\theta_1, \ldots$. The

25   values of the incremental rotor angles are averaged to obtain the final incremental rotor angle $\Delta\theta$. The knowledge of the final incremental rotor angle $\Delta\theta$ is used to control the motor.

In the first aspect, the value of the estimated

30   rotor position θ_{cal} obtained from comparison result

typically equals the reference rotor angle θ_r. However, the estimated rotor position θ_{cal} may be further modified with an incremental angle ϕ corresponding to the reference rotor angle θ_r to obtain a more accurate estimated rotor

5   position. This idea can also be extended to the second aspect.

          In a third aspect of the invention, provided is a discrete rotor position estimation method for a synchronized reluctance motor. According to the method, a

10   d.c.-link voltage V_{dc} and a phase current I_{ph} are sensed. A flux-linkage λ_{ph} of an active phase is calculated from the sensed d.c.-link voltage V_{dc} and the sensed phase current I_{ph}. The calculated flux-linkage λ_{ph} is compared with a reference flux-linkage λ_r. The reference flux-

15   linkage λ_r corresponds to a reference angle θ_r which lies between angles corresponding to aligned rotor position and non-aligned rotor position of the synchronized reluctance motor. An estimated rotor position θ_{cal} is calculated from the calculated flux-linkage λ_{ph} using either one of the

20   inductance model or the flux linkage model of the active phase, only once when the calculated flux-linkage λ_{ph} is greater than the reference flux-linkage λ_r. In this case, the ideal instant to estimate the rotor position may be at one PWM interrupt before the next phase is turned ON. From

25   the estimated rotor position θ_{cal}, incremental rotor angle $\Delta\theta$ for every PWM interrupt is calculated. The knowledge of $\Delta\theta$ is finally used to control the motor.
The above idea can be also extended to estimate the rotor position either twice or thrice such as θ_{cal1}, ... from the

30   calculated flux-linkage λ_{ph} using either one of the

inductance model or the flux linkage model of the active
phase, at every consecutive PWM interrupt when the
calculated flux-linkage λ_{ph} is greater than the reference
flux-linkage λ_r. From the estimated rotor positions $\theta_{cal1}, \ldots$

5   incremental rotor angles for every PWM interrupt are also
calculated twice or thrice such as $\Delta\theta_1, \ldots$. The values of
the incremental rotor angles are averaged to obtain the
final incremental rotor angle $\Delta\theta$. The knowledge of the
final incremental rotor angle $\Delta\theta$ is used to control the

10   motor.

In a fourth aspect of the invention, provided is
a discrete rotor position estimation method for a
synchronized reluctance motor. According to the method, a
phase inductance of the synchronized reluctance motor is

15   detected. A minimum region of the phase inductance during
turn-on of an active phase is identified. An approximate
rotor position θ_{app} is determined from the identified
minimum inductance region. From the approximate rotor
position θ_{app}, incremental rotor angle $\Delta\theta$ for every PWM

20   interrupt is calculated. The knowledge of $\Delta\theta$ is finally
used to control the motor.

In a fifth aspect of the invention, provided is a control
method of a synchronized reluctance motor and a technique
to obtain the incremental rotor angle $\Delta\theta$ for every PWM

25   interrupt. According to the control method, the estimated
rotor position θ_{cal} is obtained by the previously described
estimation methods. An absolute rotor position θ_{abs} is
calculated from the estimated rotor position θ_{cal} by adding
a stroke angle of the motor. The incremental rotor angle

30     $\Delta\theta$ is determined by processing an error between the

absolute rotor position θ_{abs} and a finally estimated rotor position θ_{est} through either one of a proportional-integral (PI) control and a proportional control. The finally estimated rotor position θ_{est} in every predetermined period

5   is generated by adding the incremental rotor angle $\Delta\theta$ to the finally estimated rotor position θ_{est} in the previous cycle. Turn-on and turn-off angles of each phase is controlled based on the finally estimated rotor position θ_{est}.

10   In a sixth aspect of the invention, provided is a control method of a synchronized reluctance motor and a technique to obtain the incremental rotor angle $\Delta\theta$ for every PWM interrupt. According to the control method, an incremental rotor angle $\Delta\theta$ is calculated by counting the

15   number of PWM interrupts between two consecutive instants when the estimated rotor position θ_{cal} is obtained by the estimation method according to the invention. Delays to turn-off an active phase and turn-on the next phase is generated, in which the delays are normally defined with

20   respect to the reference rotor position θ_r. The delays are adjusted with the estimated rotor position θ_{cal} to turn-off the active phase and turn-on the next phase. A turn-on angle θ_{on} and a turn-off angle θ_{off} of each phase of the motor are controlled based on the adjusted delays

25   decided by the incremental rotor angle $\Delta\theta$.

In the above control methods, a speed ω of the motor may be calculated from the incremental rotor angle $\Delta\theta$ in a relatively slower timer interrupt compared to a PWM interrupt, and a turn-on angle θ_{on} and a turn-off angle θ_{off}

30   of each phase of the motor may be varied continuously

based on the speed ω and the torque demand of the motor.

In a seventh aspect of the invention, provided is a control method of a synchronized reluctance motor. According to the control method, a peak of a phase current and a negative change rate of phase current in each phase are continuously monitored. Turn-off angle is kept fixed and turn-on angle is advanced so that a pre-determined peak phase current and a negative rate of change of phase current corresponding to the maximum torque are achieved.

In the control method, instead of monitoring the negative rate of change of phase current, a lead angle ϕ between the peak current and the peak flux in each phase may be detected to judge the maximum torque at the rated speed condition.

Brief Description of Drawings

Fig. 1 shows the cross-sectional view of a switched reluctance motor.

Fig. 2 shows the per phase flux-linkage / current characteristics with respect to rotor position of a switched reluctance motor.

Fig. 3 shows the per phase inductance / current characteristics with respect to rotor position of a switched reluctance motor.

Fig. 4 shows a typical inverter circuit for the switched reluctance motor.

Fig. 5 shows the region where the reference rotor position θ_r corresponding to the reference flux-linkage λ_r in rotor position estimation technique I can be defined

Fig. 6 shows the instant at which estimation of

the exact rotor position θ_{cal} is carried out for the rotor position estimation technique II.

Fig. 7 shows the instant at which the measured actual phase current I_{ph} and the estimated phase current Is are compared in the discrete rotor position estimation technique III.

Fig. 8 shows the incremental change of per phase flux-linkage λ_{ph} in every PWM interrupt of the processor in accordance with the first exemplary embodiment.

Fig. 9 shows the block diagram for closed loop control using control strategy I in accordance with the first exemplary embodiment.

Fig. 10 shows the turn-on and the turn-off delay with respect to the reference rotor position θ_r in accordance with the second exemplary embodiment.

Fig. 11 shows the block diagram for closed loop control using control strategy II in accordance with the second exemplary embodiment.

Fig. 12A shows two reference flux linkages corresponding to two reference rotor positions in accordance with the third exemplary embodiment.

Fig. 12B shows three reference flux linkages corresponding to three reference rotor positions in accordance with the third exemplary embodiment.

Fig. 13 shows the block diagram for closed loop control using control strategy I in accordance with the third exemplary embodiment.

Fig. 14 shows the block diagram using the discrete rotor position estimation technique III and the control strategy I.

Fig. 15 shows typical per phase current waveforms and per phase flux waveform at the rated speed and the maximum torque condition in accordance with the fifth exemplary embodiment.

5   Fig. 16 shows the peak current detector block introduced in the closed loop block diagram using discrete rotor position technique I and control strategy I.

Figs. 17A through 17G show examples of applications using the proposed rotor position estimation

10   schemes and control strategies for the switched reluctance motor.

Best mode for carrying out the Invention

The present invention relates to the closed loop

15   control techniques of switched reluctance motors (SRM) without a shaft position sensor where the discrete rotor position estimation techniques are followed. Three typical discrete rotor estimation techniques I, II and III are described briefly below.

20

(Discrete Rotor Position Estimation Technique I)

In the discrete rotor position estimation technique I, the per phase flux-linkage at rotor position θ_r as shown in Fig. 5 is chosen as a reference flux-

25   linkage λ_r and is expressed as a function of phase current I_{ph} by a simple polynomial expression such as,

$$\lambda_r = \Sigma A_n I_{ph}{}^n \qquad (3).$$

θ_r is defined as any rotor position near the mid-position between the aligned and the non-aligned rotor

30   position which is shown by the shaded area in Fig. 5.

Mathematically θ_r can be expressed as,

$$\theta_r = \theta_m \pm \alpha \qquad (4)$$

where, θ_m is the exact middle position between the aligned and the non-aligned rotor position and α is the deviation angle whose maximum value α_{max} is equal to 30° electrical from the middle position. θ_m and α_{max} are expressed in mechanical degrees as follows,

$$\theta_m = (\theta_a + \theta_n)/2 \qquad (5)$$

and

$$\alpha_{max} = 30°/P \qquad (6).$$

θ_a is defined as the aligned position i.e. when a rotor pole aligns with a stator pole. θ_n is defined as the non-aligned position i.e. when a rotor pole is in between two stator poles as shown in Fig. 1. The value of n in the polynomial equation (3) is dependent on the magnetic design of a particular motor. The reason why the near middle position θ_m is selected as a reference position is that the flux linkage varies with the position θ in a high changing rate in vicinity of the middle position θ_m and thus the position θ with respect to the flux linkage can be easily determined.

Use of fast processors will enable to calculate polynomial expressions easily. The reference flux-linkage value λ_r with respect to phase current I_{ph} is pre-determined off-line by locking the rotor at position θ_r and carrying out standard experiments. During motor rotation, the flux-linkage per phase λ_{ph} is calculated on-line at every PWM interrupt or at every half-cycle PWM interrupt of the processor by sensing the d.c.-link voltage V_{dc} and the phase current I_{ph} of the inverter circuit as

shown in Fig. 4. It is given by,

$$\lambda_{ph} = (V_{dc} - I_{ph}*R - V_p)*dt \qquad (7)$$

where R is the phase resistance of the SRM in the high
frequency mode and V_p is the voltage drop across the power

5   devices. An average value of mutual flux (maximum 10% of
the maximum per phase flux) should be considered for more
accurate calculation of the per phase flux-linkage λ_{ph}.
The on-line calculated per phase flux-linkage λ_{ph} is
continuously compared with the reference flux-linkage λ_r

10   in the processor. At the instant when λ_{ph} is equal to λ_r,
the absolute discrete rotor position θ_{abs} in mechanical
degrees is determined or estimated as given below

$$\theta_{abs} = K*360°/N + \theta_{cal} \qquad (8)$$

where, $\theta_{cal} = \theta_r$ and K = (N-1).

15   From the information of θ_{abs}, closed loop control
techniques are generated to rotate the motor.

(Discrete Rotor Position Estimation Technique II)

            In the discrete rotor position estimation

20   technique II, similarly the d.c.-link voltage V_{dc} and the
phase current I_{ph} are sensed and the flux-linkage per phase
λ_{ph} is calculated on-line from equation (7) at every PWM
interrupt of the processor. From the knowledge of λ_{ph}, the
exact rotor position θ_{cal} is calculated only once either

25   from the flux-linkage model or the inductance model of the
active phase when the calculated flux-linkage λ_{ph} is
greater than the reference flux-linkage λ_r. The ideal
instant to estimate the rotor position may be at one PWM
interrupt before the next phase is turned ON. This is shown

30   in Fig. 6. This concept is ideal when the turn-on angle is

advanced at the rated speed for high torque applications. The absolute discrete rotor position θ_{abs} in one mechanical cycle is always calculated from the information of the exact rotor position θ_{cal} and is expressed by equation (8).

5   From the knowledge of θ_{abs} control techniques are again generated to rotate the motor in closed loop.

(Discrete Rotor Position Estimation Technique III)

        The discrete rotor position estimation technique
10   III is very simple and does not involve any exact rotor position estimation θ_{cal} as described in techniques I and II.   This scheme includes identifying the minimum inductance region during the turn-on of a new phase so that the turn-on angle always lies in the minimum inductance
15   region.   This has been made possible by comparing the measured actual phase current I_{ph} with an estimated phase current I_s after a finite time interval t_0 from the turn-on of a new phase as shown in Fig. 7.   The estimated current I_s is given by,

20             $$I_s = (V_{dc} - I_{ph} R - V_p) * t_0 / L_{min} \quad (9).$$

        The value of minimum inductance L_{min} is constant for a particular motor and does not vary with the phase current I_{ph} as shown in Fig. 3.   The minimum inductance region normally extends to few degrees and is typically
25   dependent on the magnetic design of the motor.   Hence, the approximate rotor position θ_{app} is known when the minimum inductance region is identified.   Therefore, instead of exact rotor position θ_{cal}, the knowledge of approximate rotor position θ_{app} is used to calculate the absolute
30   discrete rotor position θ_{abs} and rotate the motor in closed

loop.

In the discrete rotor position estimation techniques of I, II and III, a rotor position estimation error less than one mechanical degree is difficult to achieve. A small error in rotor position estimation at a particular speed can decrease the motoring torque i.e. the motor performance to a large extent. Hence, optimum control techniques are invented to minimize rotor position estimation error so that high performance is always achieved for the sensorless drive of SRM's. In addition, control techniques are invented to obtain the high torque even if there is a rotor position estimation error.

(First exemplary embodiment)

In the discrete rotor position estimation technique I, the per phase flux-linkage λ_{ph} is calculated on-line in every PWM interrupt or every PWM half-cycle interrupt by sensing the d.c.-link voltage V_{dc} and the phase current I_{ph} and is defined by the equation (7) above. The incremental change of per phase flux-linkage λ_{ph} at the valley of every PWM interrupt of the processor is shown in Fig. 8 where dt is the time interval between two PWM interrupts and θ_{on} is the turn-on angle and θ_{off} is the turn-off angle. Any digital technique cannot exactly define the instant when the per phase flux-linkage λ_{ph} is equal to the reference flux-linkage λ_r. In this embodiment, the processor within a PWM interrupt identifies an instant when the per phase flux-linkage λ_{ph} is greater than λ_r as shown in Fig. 8. Therefore, for more accurate rotor position estimation, θ_{abs} of equation (8) in the

discrete rotor position technique I has to be modified.   In
the simplest technique, the modification of θ_{abs} in
equation (8) can be done by adding an angle ϕ such that,

$$\theta_{abs} = K*360o/N + (\theta_r + \phi)   (10).$$

5   The combined angle ($\theta_r + \phi$) now defines the
exact rotor position θ_{cal}.   The angle ϕ can be obtained
from a pre-determined two dimensional look-up table in
which $\Delta\lambda$ and phase current I_{ph} are the variable parameters.
$\Delta\lambda$ is defined as the difference between λ_{ph} and λ_r.

10   Alternatively, by the discrete rotor position
estimation technique II, exact rotor position θ_{cal} should
be calculated at the instant when the per phase flux-
linkage λ_{ph} is greater than the reference flux-linkage λ_r.
Exact rotor position estimation can be carried out by
15   analytical methods based on the flux-linkage model or the
inductance model of the SRM.   The absolute discrete rotor
position θ_{abs} in one mechanical cycle is always calculated
from the information of the exact rotor position θ_{cal} and
is expressed by equation (8).   The value of θ_{abs} which is
20   defined in mechanical degrees in equation (8) can also be
expressed in electrical degrees by considering the number
of rotor poles (P).

          Fig. 9A shows an entire block diagram of a
driving apparatus of SRM for achieving the closed loop
25   control with control strategy I according to the invention.

          The SRM 1 is driven by the inverter ~~2~~ (2) which
includes switching devices T1 to T6 as shown in Fig. 4.
The gate drive circuit ~~22~~ (22) generates pulse width
modulated base drive signals to turn-on and turn-off the
30   switching devices T1 to T6 and generate a phase current.

PWM control is done by the block ~~21~~ (21) with PWM interrupt
timer. Turn-on angle θ_{on} and turn-off angle θ_{off} are
calculated by the block ~~19~~ (19). The turn-on and turn-off
controller ~~18~~ (18) determines the turn-on and the turn-off

5   timings of each phase based on the calculated turn-on and
turn-off angles θ_{on} and θ_{off} and the estimated rotor
position θ_{est}.

In the control strategy I involving the discrete
rotor position estimation technique I, phase current I_{ph}

10   and d.c.-link voltage V_{dc} are sensed to calculate the per
phase flux-linkage λ_{ph} (by block (11)). The flux-linkage
λ_{ph} is compared with the reference flux-linkage λ_r (by
block (11a)). Referring to the comparison result, when the
phase flux-linkage λ_{ph} is greater than the reference flux-

15   linkage λ_r, the rotor position θ_{cal} is once calculated (by
block (12)). The absolute discrete rotor position θ_{abs} is
calculated (by block (13)) and continuous rotor position θ_{est} is estimated from the absolute discrete rotor position
θ_{abs} in the following way (by block (14)). At the instant

20   when the exact rotor position θ_{cal} is calculated (by block
(12)), the error between the absolute rotor position θ_{abs}
and the estimated rotor position θ_{est} is calculated and
processed in a proportional-integral (PI) control method
(by block (15)) to give the incremental rotor angle $\Delta\theta$ in

25   every PWM interrupt. $\Delta\theta$ is expressed as

$$\Delta\theta = Kp * \theta_{err} + \Sigma KI * \theta_{err} \qquad (11)$$

where, $\theta_{err} = (\theta_{abs} - \theta_{est})$. At the start of rotor position
estimation, θ_{est} is initialized as θ_{abs} and then calculated
as follows

30    $$\theta_{est}(n) = \theta_{est}(n-1) + \Delta\theta \qquad (12).$$

Instead of a proportional-integral control method, a proportional control method can also be used.

The turn-on angle θ_{on} and turn-off angle θ_{off} of each phase is controlled by θ_{est} which can be expressed either in mechanical or electrical degrees (by block (18)). For achieving the turn-on and the turn-off of each phase at any point in between two PWM interrupts, a very fast timer interrupt compared to the PWM interrupt has been defined.

The speed (ω) of the motor is calculated (by block (16)) from the incremental rotor angle $\Delta\theta$ in a relatively slow timer interrupt compared to the PWM interrupt. It is given by,

$$\omega = \Delta\theta * fs \qquad (13)$$

where, fs is the PWM carrier frequency. The speed ripple is filtered by processing it through a low-pass filter (LPF) ~~17~~(17). If the per phase flux-linkage λ_{ph} is calculated in every half-cycle PWM interrupt, the incremental rotor displacement $\Delta\theta$ has to be defined for every half-cycle PWM interrupt and the speed (ω) of the motor is expressed as,

$$\omega = 2* \Delta\theta * fs \qquad (14).$$

The turn-on and the turn-off angle together with the phase voltage reference ($V_{ph}*$) or phase current reference ($I_{ph}*$) are continuously calculated within the processor and are varied depending on the speed (ω) and the torque demand of the motor (by blocks (18)-(20)). In closed loop control, the difference between the commanded speed ($\omega*$) and the calculated speed (ω) is processed through a proportional-integral (PI) control block ~~20~~ (20) to generate either the phase voltage reference ($V_{ph}*$) or

the phase current reference (I_{ph}*). Either the phase voltage reference (V_{ph}*) is compared with the PWM carrier waveform or the phase current reference (I_{ph}*) is compared with the original phase current I_{ph} to finally generate the

5   pulse-width modulated (PWM) base drive waveforms (by block (21)).

A gate drive circuit ~~22~~ (22) controls an inverter ~~2~~ (2) using the pulse-width modulated (PWM) waveforms and turn-on and turn-off angles from the turn-on and turn-off

10   controller ~~18~~ (18) to drive the SRM ~~1~~ (1).

(Second exemplary embodiment)

Fig. 11 shows an entire block diagram of a driving apparatus of SRM for achieving the closed loop

15   control with a control strategy II according to the invention.

In the control strategy II involving the discrete rotor position estimation technique I, instead of calculating θ_{est} continuously to control the turn-on and

20   turn-off angle the incremental rotor displacement $\Delta\theta$ in every PWM interrupt can be calculated in an alternative way which can generate appropriate delays to turn-off the active phase and turn-on the next phase. The speed (ω) of the motor is also calculated from the incremental rotor

25   angle $\Delta\theta$ (by block (16)), and the closed loop control can be executed in the similar manner described in the first exemplary embodiment. In the control strategy II, $\Delta\theta$ can be calculated by counting the number of PWM interrupts between two consecutive instants when the per phase flux-

30   linkage λ_{ph} is greater than the reference flux-linkage λ_r

(by blocks (13a) and (15a)), that is, between the instants
of $\theta_{cal}(n)$ and $\theta_{cal}(n-1)$. It is given by,

$\Delta\theta$ = stroke angle(S)/number of PWM interrupts   (15).

In equation (15), $\Delta\theta$ is expressed in mechanical
degrees but also can be expressed in electrical degrees by
considering the number of rotor poles.   The exact rotor
position θ_{cal} at the instant when the per phase flux-linkage
λ_{ph} is greater than the reference flux-linkage λ_r is
calculated in a similar way as described in the first
exemplary embodiment.   Calculation of θ_{cal} helps to adjust
the turn-on and the turn-off delay and this is explained as
follows.   The turn-off delay ($x°$) of the active phase and
the turn-on delay ($y°$) of the next phase are always defined
with respect to the reference rotor position θ_r as shown
in Fig. 10 for a specific speed (ω) of the motor (by block
(18b)).   The difference ϕ between the exact rotor position
θ_{cal} and the reference rotor position θ_r is always estimated
so that the turn-off delay ($x°$) of the active phase and the
turn-on delay ($y°$) of the next phase are modified as $X_1°$
and $Y_1°$, respectively (by block (18a)).   The modified turn-
off delay ($X_1°$) and the turn-on delay ($Y_1°$) are now
redefined with respect to the exact rotor position θ_{cal},
and can be expressed as,

$X_1° = x° - \phi$ (16) and

$Y_1° = y° - \phi$ (17).

For achieving the turn-on and the turn-off of
each phase at any point in between two PWM interrupts, a
very fast timer interrupt compared to the PWM interrupt has
been defined as explained in the first exemplary embodiment.
In addition, if the per phase flux-linkage λ_{ph} is calculated

in every half-cycle PWM interrupt, the incremental rotor displacement $\Delta\theta$ has to be defined for every half-cycle PWM interrupt.

The above control technique will help to achieve both high performance as well as maximum torque of the motor at the rated speed.

(Third exemplary embodiment)

The discrete rotor position estimation technique I described in the first and the second exemplary embodiment includes comparing the per phase flux-linkage λ_{ph} with only one reference flux-linkage λ_r defined for a rotor position θ_r. Hence, in both cases, exact rotor position θ_{cal} and incremental rotor angle $\Delta\theta$ are calculated only once at the instant when the per phase flux-linkage λ_{ph} is greater than the reference flux-linkage λ_r during the conduction of each active phase. The discrete rotor position estimation technique I can be extended by defining two reference flux-linkages λ_{r1} and λ_{r2} at rotor positions θ_{r1} and θ_{r2} respectively as shown in Fig. 12A or three reference flux-linkages λ_{r1}, λ_{r2} and λ_{r3} at rotor positions θ_{r1}, θ_{r2} and θ_{r3} respectively as shown in Fig.12B by polynomial expression in phase current I_{ph}. All these rotor positions lie near the mid-position θ_m with a deviation angle α_{max} of 30°. Hence, the on-line estimated per phase flux-linkage λ_{ph} is compared with two or three reference flux-linkages and accordingly two exact rotor positions (θ_{cal1} and θ_{cal2}) or three exact rotor positions (θ_{cal1}, θ_{cal2} and θ_{cal3}) are calculated during the active conduction of each phase based on the comparison result.

Alternatively, the discrete rotor position estimation technique II can be extended by calculating the exact rotor positions either twice (θ_{cal1} and θ_{cal2}) or thrice (θ_{cal1}, θ_{cal2} and θ_{cal3}) from the calculated flux-linkage λ_{ph} by using either one of the inductance model or the flux linkage model of the active phase, at every consecutive PWM interrupt when the calculated flux-linkage λ_{ph} is greater than the reference flux-linkage λ_r corresponding to reference rotor angle θ_r.

Now from θ_{cal1},... by following either the control strategy I or the control strategy II, incremental rotor angle for every PWM interrupt is also calculated twice ($\Delta\theta_1$ and $\Delta\theta_2$) or thrice ($\Delta\theta_1$, $\Delta\theta_2$ and $\Delta\theta_3$) during the active conduction of each phase. The final incremental rotor angle $\Delta\theta$ for every PWM interrupt is the average of all the calculated incremental rotor angle. The final incremental rotor angle $\Delta\theta$ is used for the estimation of rotor position θ_{est} in control strategy I or providing the necessary delay for the turn-on or the turn-off a phase as described in control strategy II. The speed (ω) of the motor is also calculated from the final incremental rotor angle $\Delta\theta$. A control block diagram using the control strategy I and calculating the incremental rotor angle for every PWM interrupt twice is shown in Fig. 13. In this figure, absolute discrete rotor positions θ_{abs1} and θ_{abs2} are calculated, respectively (by blocks (13a) and (13b)). Then the incremental rotor angles ($\Delta\theta_1$ and $\Delta\theta_2$) are calculated respectively and averaged to obtain $\Delta\theta$ (by blocks (15a) to (15c)). A control block diagram for calculating the incremental rotor angle thrice

can be similarly obtained.

(Fourth exemplary embodiment)

In the discrete rotor position estimation

5   technique II, the d.c-link voltage V_{dc} and the phase

current I_{ph} are sensed and the flux-linkage per phase λ_{ph}

is calculated on-line from equation (7) at every PWM

interrupt of the processor. From the knowledge of λ_{ph}, the

exact rotor position θ_{cal} is calculated either from the

10   flux-linkage model or the inductance model of the active

phase at one PWM interrupt before the next phase is turned

ON. This is shown in Fig.6. This concept is ideal when

the turn-on angle is advanced at the rated speed for high

torque applications. The absolute discrete rotor position

15   θ_{abs} in one mechanical cycle is always calculated from the

information of the exact rotor position θ_{cal} and is

expressed by equation (7).

From the knowledge of θ_{abs}, control strategies I

and II are again applied in the discrete rotor position

20   estimation technique II to rotate the motor in closed loop.

(Fifth exemplary embodiment)

There exist many industrial applications which

require motor operation only at the rated speed and the

25   maximum torque condition instead of the continuous variable

speed-torque operation. For such applications, the

discrete rotor position estimation technique III is

proposed.

Fig. 14 shows an entire block diagram of driving

30   apparatus of SRM using the discrete rotor position

estimation technique III and the control strategy I.   It should be noted that identical block diagram can be drawn for the discrete rotor position estimation technique III and the control strategy II.

5   The discrete rotor position technique III is very simple and does not involve any exact rotor position estimation θ_{cal} from the flux-linkage model or the inductance model as described in techniques I and II.   The discrete rotor position technique III includes identifying

10   the minimum inductance region during the turn-on of a new phase so that the turn-on angle always lies in the minimum inductance region.   This has been made possible by comparing the measured actual phase current I_{ph} with the estimated phase current I_s after a finite time interval t_0

15   from the turn-on of a new phase as shown in Fig. 7.   The estimated current I_s is calculated from equation (9) in the block ~~11a~~(11a).

The locking of a particular phase and open loop starting technique with a forced drive always ensures that

20   the turn-on angle is initially synchronized with the minimum inductance region.   Coinciding the turn-on angle always with the minimum inductance region also guarantees perfect synchronous operation of the motor.   Initially the turn-on angle can be anywhere between θ_1 and θ_2 as shown

25   in Fig. 7.   However, the turn-on angle is initialized with θ_0 which is the middle position between θ_1 and θ_2. Therefore, the approximate rotor position θ_{app} at t_0 can be easily calculated.   In this embodiment, instead of exact rotor position θ_{cal}, approximate rotor position θ_{app} is

30   obtained (in block (12a)), and the knowledge of approximate

rotor position θ_{app} is used to calculate the absolute discrete rotor position θ_{abs} (in block (13b)) and rotate the motor in closed loop by applying either control strategy I or II.

5

(Sixth exemplary embodiment)

In the discrete rotor position estimation techniques of I, II and III, a rotor position estimation error less than one mechanical degree is difficult to

10   achieve. At the rated speed and the maximum torque condition, the SRM operates in the single pulse mode. Typical per phase current waveforms with a variable turn-on angle and a fixed turn-off angle at the rated speed are shown in Fig. 15. The per phase flux waveform

15   corresponding to the rated speed and the maximum torque condition is also shown in Fig. 15. From Fig. 15 it can be easily understood that the peak phase current (I_{ph}(peak)) which is immediately followed by the negative rate of change of phase current has to be continuously monitored if

20   a specific application requires maximum torque operation at the rated speed. Therefore, keeping the turn-off angle fixed, the turn-on angle is advanced until the pre-determined reference peak phase current (I_p*) and a reference negative rate of change of phase current

25   (dI_{ph}*/dt) corresponding to the maximum torque are achieved. Instead of monitoring the negative rate of change of phase current, the lead angle (ϕ) between the peak current (I_{ph}(peak)) and the peak flux (λ_{ph}(peak)) can be also monitored to judge the maximum torque at the rated speed

30   condition. As shown in Fig. 16, the peak current detector

block ~~30~~ (30) is introduced in the closed loop block
diagram using discrete rotor position technique I and
control strategy I.

5   (Industrial Applicability)

          The proposed discrete position sensorless
estimation schemes for the switched reluctance motors are
ideal for several automotive applications such as the
compressor drive in car air-conditioners for a conventional
10   gasoline vehicle, an electric vehicle and a hybrid electric
vehicle.

          Examples of applications using the proposed rotor
position estimation schemes and control strategies for the
switched reluctance motor are shown in Figs. 17A through
15   17G.

          Fig. 17A is an explanatory diagram of an
application of the invention to an electric vehicle. The
drive unit ~~100~~ (100) includes a switched reluctance motor
~~1~~(1), an inverter ~~2~~(2), a current sensor ~~5~~(5), and SRM
20   control system ~~50~~(50). Typically it means that the SRM
control system ~~50~~ (50) has the same structure as that shown
in Fig. 9, 11, 13, 14 or 16 except for the SRM ~~1~~(1). The
drive unit ~~100~~ (100) is coupled with a main ECU (Electrical
Control Unit) ~~61~~ (61) and a gear ~~63~~(63).

25   Fig. 17B also shows an application of the
invention to a hybrid electric vehicle. The engine ~~65~~ (65)
is further provided to the structure of the Fig. 17B. Fig.
17C shows an application of the invention to a water pump.

          Fig. 17D shows an application of the invention to
30   a compressor drive of an air conditioner or a refrigerator.

The drive unit ~~100~~ (100) drives a compressor ~~70~~ (70) which compresses refrigerant circulating in a refrigeration cycle ~~71~~ (71).

Fig. 17E shows an application of the invention to a machine tool.   Fig. 17F shows an application of the invention to a cleaner.   Fig. 17G shows an application of the invention to a fan device.

However,   the   proposed   discrete   position sensorless estimation techniques can be extended to any industrial applications involving these types of motors where the speed response required is low.

Although the present invention has been described in connection with specified embodiments thereof, many other   modifications,   corrections   and   applications   are apparent to those skilled in the art.   Therefore, the present invention is not limited by the disclosure provided herein but limited only to the scope of the appended claims.